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Isomura, et al. **Attorney Docket No.**: TNG-3-US

U.S. Patent No.: 6,762,658 B1 **Group Art Unit**: 2817

Date Issued: July 13, 2004 **Examiner:** Takaoka, Dean O.

For: DIELECTRIC RESONATOR AND DIELECTRIC FILTER

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

SUBMISSION OF POSSIBLY RELATED PRIOR ART

Sir:

For the possible benefit of anyone subsequently evaluating the scope and/or validity of the above patent, it is respectfully requested that the Supplementary European Search Report dated April 5, 2004 and European Office Action dated July 5, 2004 (copies attached), filed in the European counterpart of U.S. Patent Application No. 09/807,819, along with the following reference cited in the Supplementary European Search Report, be placed in the file:

> Awai, et al., "A Dual Mode Dielectric Waveguide Resonator and its Application to Bandpass Filters", IEICE Transactions on Electronics, Institute of Electronics Information and Comm. Eng. Tokyo, JP, Vol. E78-C, No. 8, August 1, 1995, pp. 1018-1025.

The undersigned has not reviewed the teachings of the European Office Action, the Supplementary European Search Report, nor the cited reference in detail and thus makes no representations concerning their relevancy or materiality.

Respectfully submitted,

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Anmelder/Applicant/Demandeur/Patentinhaber/Proprietor/Titulaire
Kabushiki Kaisha Tokin, et al

COMMUNICATION

The European Patent Office herewith transmits as an enclosure the European search report for the above-mentioned European patent application.

If applicable, copies of the documents cited in the European search report are attached.

☒ Additional set(s) of copies of the documents cited in the European search report is (are) enclosed as well.

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REFUND OF THE SEARCH FEE

If applicable under Article 10 Rules relating to fees, a separate communication from the Receiving Section on the refund of the search fee will be sent later.

European Patent Office

SUPPLEMENTARY EUROPEAN SEARCH REPORT

Application Number
EP 00 95 3537

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X	AWAI I ET AL: "A DUAL MODE DIELECTRIC WAVEGUIDE RESONATOR AND ITS APPLICATION TO BANDPASS FILTERS" IEICE TRANSACTIONS ON ELECTRONICS, INSTITUTE OF ELECTRONICS INFORMATION AND COMM. ENG. TOKYO, JP, vol. E78-C, no. 8, 1 August 1995 (1995-08-01), pages 1018-1025, XP000536085 ISSN: 0916-8524 * page 1021, left-hand column, lines 21-27; figure 9a * -----	6,8	H01P1/20 H01P1/208 H01P7/10

TECHNICAL FIELDS SEARCHED (Int.Cl.7)

H01P

The supplementary search report has been based on the last set of claims valid and available at the start of the search.

Place of search	Date of completion of the search	Examiner
The Hague	29 March 2004	Den Otter, A

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
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2334c IEICE Transactions on Electronics
E78-C(1995) August, No. 8, Tokyo, JP

PAPER Special Issue on Microwave and Millimeter-Wave Technology

A Dual Mode Dielectric Waveguide Resonator and Its Application to Bandpass Filters

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p. 1018 – 1025 : 8

SUMMARY The fundamental TE_{10} mode in a rectangular waveguide of a square cross section is degenerate with TE_{01} mode. A quarter wavelength resonator made of a dielectric square waveguide is, therefore, applied for a small-sized bandpass filter, just like dual mode filters for base stations in the mobile communication. In this paper, the methods to couple the two modes are first studied, including cutting a corner of the resonator and adding some metal electrodes on its end face. Both methods help to flow the *rf* current of the odd mode at the corner, resulting in decrease of the series inductance and thus increase of the resonant frequency. The coupling constant, that is proportional to the difference of the odd and even-mode's resonant frequency, can be controlled by the perturbations mentioned above. The coupling to the external circuit is adjusted by an electrode fabricated also on the end face. It is connected to a microstrip line and capacitively couples to the resonant modes. The coupling strength increases with the dimension of the electrode. The adjustment of the resonant frequency is carried out by the similar electrode on the end face and connected to the center of the side of the square cross section. The frequency decreases with the length of the electrode. The unloaded Q is measured to be of around 500 for $5\times5\times10$ mm resonator of $\varepsilon_r=93$. The optimum aspect ratio for the resonator is found in terms of the Q value. The simplest bandpass filter, i.e., a two-stage bandpass filter is designed and fabricated using $5\times5\times10$ mm resonator. It is mounted in a square hole made in a printed circuit board and excited by a microstrip line. The frequency characteristics are in good agreement with the expected values.

***key words:** dual mode, dielectric waveguide resonator, high permittivity, bandpass filter*

1. Introduction

In the mobile communication system, dual-mode circular cylindrical or rectangular dielectric resonators have significantly contributed to make a bandpass filter more compact [1]. But those resonators are basically three-dimensional and, moreover, usually encapsulated in a metal case to avoid radiation, resulting in a rather bulky configuration in spite of the multiple spatial use of a resonator. Portable stations, on the other hand, use much smaller BPFs made of coaxial or strip line resonator [2], which are basically of one-dimensional structure. They are miniaturized to the limit at the expense of insertion loss.

There could be an intermediate type of BPF which has moderate physical dimension and insertion loss. It may be realized by a circular cylindrical or rectangular resonator directly covered with metal, which eventually results in a dielectric waveguide resonator [3]. These intermediate characters would be useful for a microwave resonator or filter such as of a small base station which dose not handle too much power nor needs too small dimensions. Our investigation is on those resonators of square cross section. They have degenerate orthogonal modes to be coupled each other by adding a slight perturbation to the resonator [4].

Similar structures have already been studied in the form of a rectangular waveguide [5] or microstrip line resonator [6]. Since a dielectric material of high permittivity has strong field confinement, our resonator can enjoy a variety of coupling schemes different from the two structures mentioned above. We will show here two main coupling methods, one of which is cutting the corner of the resonator and the other adds some metallic patterns on the open end face.

The coupling to the external circuit, frequency tuning of each resonating mode and the unloaded Q of the resonator will also be investigated to design a bandpass filter. In the last section, design and fabrication of BPFs are shown based on the data in the following sections.

2. Basic Structure of the Resonator

The fundamental mode of a hollow rectangular waveguide resonator shown in Fig. 1(a) is TE_{101} mode if its dimensions a, b and c are in the order $b\leqq a\leqq c$. It is considered to be a half wavelength resonator in the c direction. When the cross section is square (Fig. 1(b)), that is $a=b$, TE_{101} mode becomes degenerate with TE_{011} mode and they constitute the lowest mode pair.

A dielectric waveguide of the same shape covered with metal film has, in principle, the same property except that the resonant frequency decreases by the dielectric constant. Thus the electric field of the degenerate TE_{101} and TE_{011} modes are respectively given by

$$e_1=\sin\left(\frac{\pi}{a}x\right)\sin\left(\frac{\pi}{c}z\right)a_y \tag{1}$$

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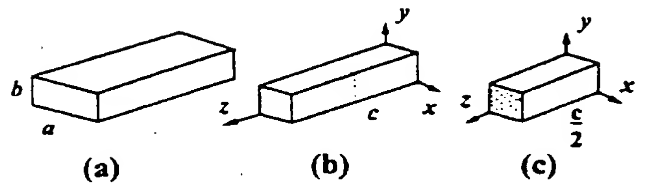

Fig. 1 Rectangular waveguide resonators.
(a) Hollow waveguide $\lambda/2$ resonator
(b) Dielectric waveguide $\lambda/2$ resonator of square cross section
(c) Dielectric waveguide $\lambda/4$ resonator of square cross section

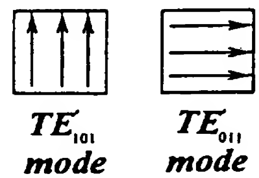

Fig. 2 Transverse electric field distribution of two degenerate fundamental modes in $\lambda/4$ dielectric resonator.

$$e_2 = \sin\left(\frac{\pi}{a}y\right)\sin\left(\frac{\pi}{c}z\right)a_x, \tag{2}$$

where a_x and a_y are a unit vector along x and y direction, respectively, according to the coordinate system shown in Fig. 1 (b). Now, one cut the resonator in the center along the dotted line in Fig. 1 (b), then a quarter wavelength resonator of Fig. 1 (c) is obtained. Due to the high permittivity, the open end face makes a magnetic wall approximately. Thus the electromagnetic field in the resonator does not change from that in the half wavelength resonator. The lowest degenerate modes in $\lambda/4$ resonator are called TE'$_{101}$ and TE'$_{011}$ whose electric fields are roughly described in Fig. 2.

3. Metal-Covered Corner Cut

The effect of a partial corner cut as shown in Fig. 3 is analyzed with the help of the variational method. If we assume E is the electric field in the dielectric resonator, the resonant frequency is given by [7]

$$\omega_r^2 = \frac{\iiint \mu^{-1}(\nabla \times E)^2 d\tau + 2\iint [(\mu^{-1}\nabla \times E) \times E]\cdot dS}{\iiint \varepsilon E^2 d\tau} \tag{3}$$

This expression is stationary, even if the tangential electric field dose not vanish on the resonator surface. Thus, we can expect a good approximation without taking care of the boundary condition for the trial function E_{tr}.

As for E_{tr}, we choose the following linear combination of TE'$_{101}$ and TE'$_{011}$ modes

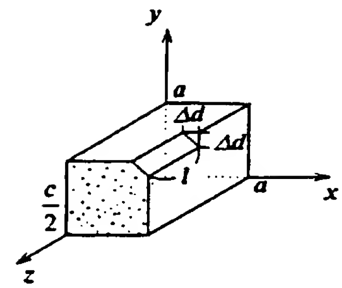

Fig. 3 Configuration of corner-cut $\lambda/4$ resonator of square cross section.

$$E_{tr} = m_1 e_1 + m_2 e_2 \tag{4}$$

The quantity e_1 and e_2 are the degenerate dominant modes in the square dielectric waveguide resonator of a quarter wavelength given in Eqs. (1) and (2). Substituting Eq. (4) into Eq. (3), we obtain

$$\omega_r^2 = \frac{\sum_{i=1}^{2}\sum_{j=1}^{2} m_i m_j N_{ij}}{\sum_{i=1}^{2}\sum_{j=1}^{2} m_i m_j^* D_{ij}} \tag{5}$$

$$N_{ij} = \iiint \mu_0^{-1}\nabla \times e_i \cdot \nabla \times e_j d\tau + 2\iint \mu_0^{-1}(\nabla \times e_i) \times e_j \cdot dS \tag{6}$$

$$D_{ij} = \iiint \varepsilon e_i \cdot e_j^* d\tau \tag{7}$$

Following the Rayleigh-Ritz procedure, we differentiate Eq. (5) with respect to m_i and obtain the secular equation by putting the determinant of the coefficient matrix of m_1 and m_2 equal to zero.

$$(N_{11} - D_{11}\omega_r^2)(N_{22} - D_{22}\omega_r^2) - (N - D\omega_r^2)^2 = 0 \tag{8}$$

where

$$\frac{N_{12} + N_{21}}{2} = N, \quad \frac{D_{12} + D_{21}}{2} = D \tag{9}$$

Now, considering the symmetry of the structure, we have

$$D_{11} = D_{22}, \; N_{11} = N_{22}, \; N_{12} = N_{21}, \; D_{12} = D_{21} = 0 \tag{10}$$

Solving Eq. (8) for ω_r^2, we get as the final result

$$\omega_r^2 = \begin{Bmatrix}\omega_e^2 \\ \omega_o^2\end{Bmatrix} = \frac{\pi^2}{\varepsilon\mu_0}\left(\frac{1}{c^2} + \frac{1}{a^2}\right) + \frac{\pi^2}{\varepsilon\mu_0 c a^4} \cdot \left\{\frac{3a}{\pi}\sin\left(\frac{\pi}{a}\Delta d\right) \pm \Delta d\right\}\left\{\frac{a}{\pi}\sin\left(\frac{\pi}{a}\Delta d\right) \mp \Delta d\right\} \cdot \left\{l + \frac{c}{2\pi}\sin\left(\frac{2\pi}{c}l\right)\right\}, \tag{11}$$

where $\pm$ signs are for the combinations of $m_1 = \pm m_2$ and correspond to the even mode angular frequency ω_e and that of odd mode ω_o, respectively. The first term

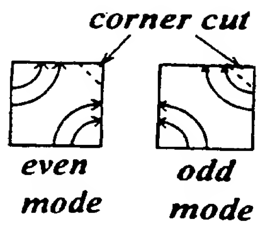

Fig. 4 Field distribution of even and odd modes.

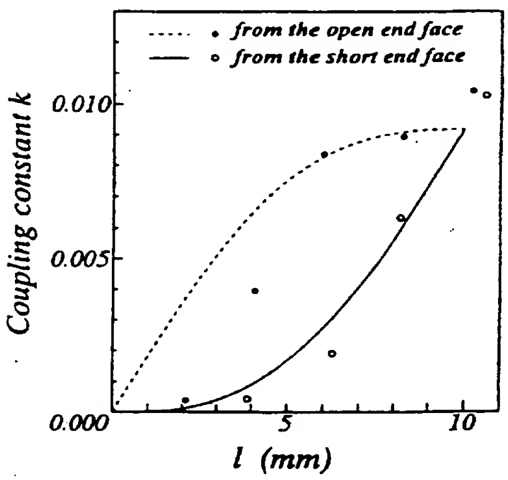

Fig. 5 Coupling constant versus corner cut length. ($a=5$ mm, $c/2=10$ mm $\Delta d=0.6$ mm $\varepsilon_r=93$)

of Eq. (11) represents the squared value of the degenerate resonant frequency of unperturbed TE'_{101} or TE'_{011} mode. The coupling constant k is calculated as

$$k=\frac{2(\omega_o^2-\omega_e^2)}{\omega_e^2+\omega_o^2} \tag{12}$$

The field distributions of the even and odd modes are described in Fig. 4 The names correspond to the symmetry in regard to the corner cut.

(a) Dependence on the Cut Length

The coupling constant naturally increases as the cut length. But the location of the partial cut along the ridge influences the coupling, because the electric field varies along the longitudinal direction (z-axis). It takes maximum at the open end face, therefore the contribution of the corner cut should be largest there, considering Eq. (5). The theoretical and experimental coupling constants are plotted in Fig. 5 for either cases when cutting is begun from the open or short end face. The dielectric material has a relative permittivity of 93, and the resonator dimension is $5\times5\times10$ mm hereafter.

The main reason of discrepancy between theory and experiment is supposed to be the difficulty in machining the corner accurately and covering it with aluminum evaporation again.

(b) Dependence on the Cut Depth

A greater depth of cutting should also increase the coupling constant. There are shown theoretical and experimental results in Fig. 6 with a good agreement.

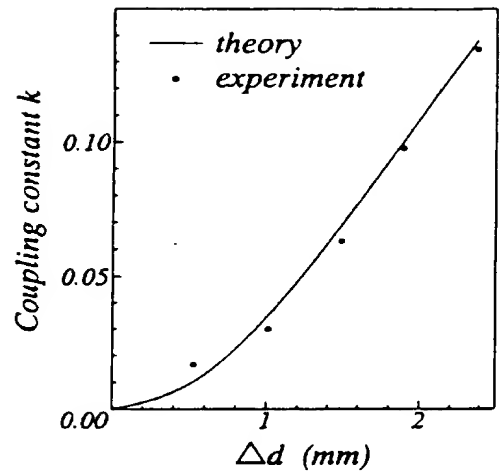

Fig. 6 Coupling constant versus corner cut depth. ($a=5$ mm, $c/2=10$ mm, $l=10$ mm $\varepsilon_r=93$)

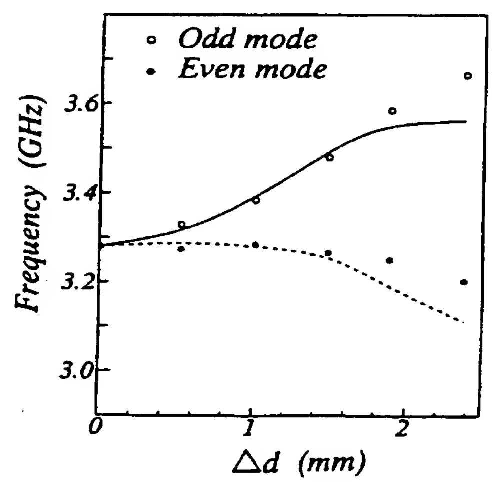

Fig. 7 Resonant frequency versus corner cut depth. ($a=5$ mm, $c/2=10$ mm, $l=10$ mm $\varepsilon_r=93$)

The original data for the even and odd mode's resonant frequency are shown in Fig. 7 together with the calculated value by Eq. (11). The theoretical result was shifted arbitrarily to fit the experimental one at $\Delta d=0$ mm. We can explain that the even mode is not affected by the corner cut, as follows. The *rf* current flowing across the unperturbed corner ($x=a$, $y=a$ in Fig. 3) can be obtained substituting Eqs. (1) and (2) into Maxwell's equation. The crossing currents at the corner for the even and odd modes are calculated as

$$I^e=0,\quad I^o=\frac{2a}{j\omega\mu\pi}\sin\left(\frac{\pi}{c}z\right) \tag{13}$$

The corner cut shortens the current path, resulting in the reduction of inductance for the odd mode. There-

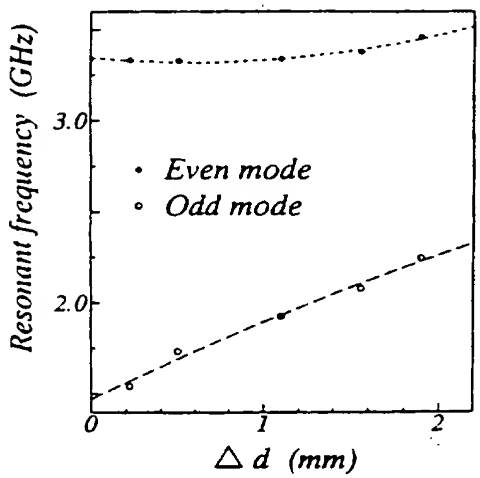

Fig. 8 Resonant frequency versus open corner cut depth. ($a=5$ mm, $c/2=10$ mm, $l=10$ mm, $\varepsilon_r=93$)

fore, the even mode is not affected at all, whereas the odd mode's frequency increases with the perturbation.

(c) Unmetalized Corner Cut

If the cut corner is not covered with metal, it will intercept the *rf* current across the corner, and thus, the field distribution of the odd mode will be influenced appreciably. That of the even mode, on the contrary, will not change because the wall current at the corner dose not exist as shown in Eq. (13).

The experimental result tells that the frequency shift of the odd mode is as much as 1 to 2 GHz, while that of the even mode is almost kept constant with the depth of cutting (Fig. 8). Considering that the coupling constant is determined by the difference of the resonant frequencies of both modes, i.e. by Eq. (12), it is too strong to create a BPF of reasonable bandwidth and also too sensitive to tune by a step-by step cutting. Therefore, we will discard this coupling scheme in spite of its simplicity and ease of adjustment after fabrication.

4. Metal Patterns on the End Face

The open end face of a $\lambda/4$ dielectric waveguide resonator can be used for several purposes. First, a shorted conductor at the corner shown in Fig. 9(a) would act as a perturbation to couple the degenerate modes. It will help flow the *rf* current across the corner, and increase the frequency of the odd mode without affecting the even mode. Secondly, a shorted metal strip at the center of a square side as shown in Fig. 9(b) will be used to tune the resonant frequency. Lastly, a metal strip connected to the external circuit will excite the resonator (Fig. 9(c)). The excitation should become stronger as the area of the strip increases.

(a) Coupling of the Degenerate Modes

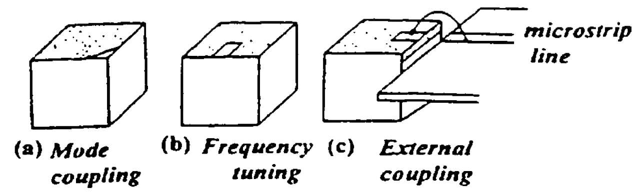

Fig. 9 Three kind of metal patterns for different purposes.

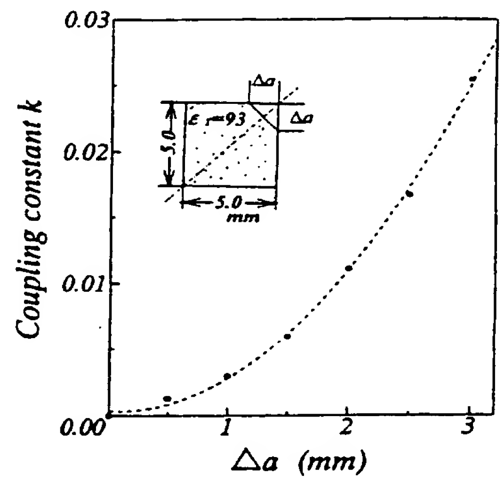

Fig. 10 Coupling constant versus dimension of the triangular metal pattern. ($c/2=10$ mm)

Considering that the *rf* current across the corner is given by Eq. (13), we can understand the metal on the end face like Fig. 9(a) will make a bypass for the current of the corner cut as described before. But it should be smaller because the current path is far narrower in this case than the corner cut shown in Fig. 3. An experimental result is given in Fig. 10. As was expected, the coupling strength is around one tenth of Fig. 6, to the same dimension of the triangular pattern as the triangular cross section in Fig. 3. The even or odd mode is named according to the symmetry to the diagonal line crossing the perturbation as was before. Thus, the names are interchanged if the perturbation is shifted to the neighboring corners, e.g. ($x=a$, $y=0$) or ($x=0$, $y=a$) in Fig. 1(c). This fact suggests that two neighboring perturbation contribute subtractively to the coupling constant. We have carried out an experiment to confirm it, as is shown in Fig. 11. The patterns M_1 and M_1' have increased the frequency of only the even mode and are kept unchanged. (The names even or odd are referred to the pattern M_2) Then, we add the pattern M_2 along the perpendicular diagonal line. It increases only the frequency of the odd mode, and thus, those of both modes intersect at $\Delta a=1.6$ mm. Considering that the coupling constant is given by Eq. (12), it becomes zero at the point, changing the sign when Δa passes the point. As a result, these patterns

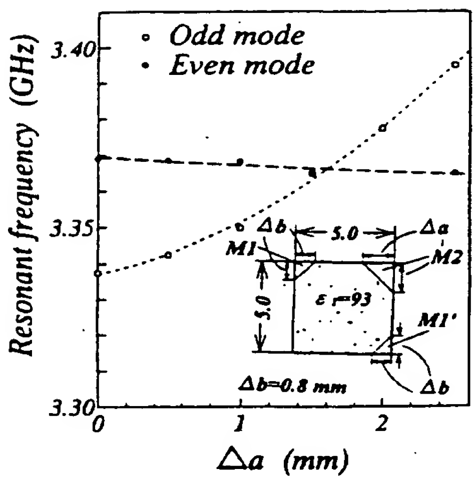

Fig. 11 Adjustment of coupling by negative perturbation. (M1 and M1' are kept constant. $c/2=10$ mm)

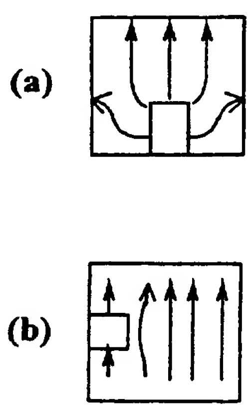

Fig. 12 Electric field distribution for two electrode configurations.
(a) parallel electrode
(b) perpendicular electrode

could be used to adjust the coupling constant after setting the main coupling by the corner cut.

(b) Frequency Tuning

The resonant frequency of both modes could be slightly different because of the unexpected deformation from a square cross section of the resonator. A metal strip connected to the side of the square end face shown in Fig. 9(b) would behave as a shunt capacitor to reduce the resonant frequency. It will affect TE'_{101} mode, but not TE'_{011} mode. The electric field of the former will be changed as shown in Fig. 12(a), whereas that of the latter will remain almost the same as in Fig. 12(b). Figure 13 describes the experimental result. The curve (b) tells the parallel electrode of 2.0×1.8 mm tunes the frequency as much as 10%, while the perpendicular one (a) has a negligibly small effect.

(c) Coupling to the External Circuit

One possible method of mounting the resonator is

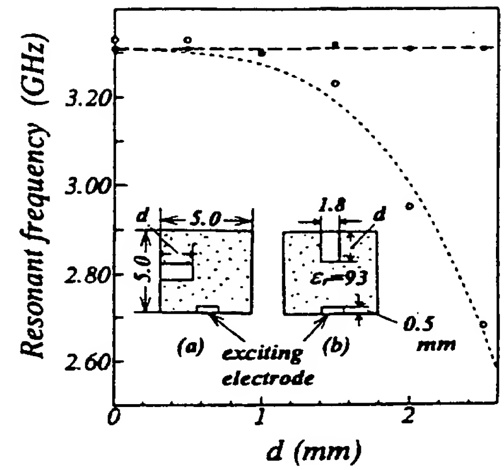

Fig. 13 Resonant frequency tuning of TE_{101} mode. ($c/2=10$ mm)

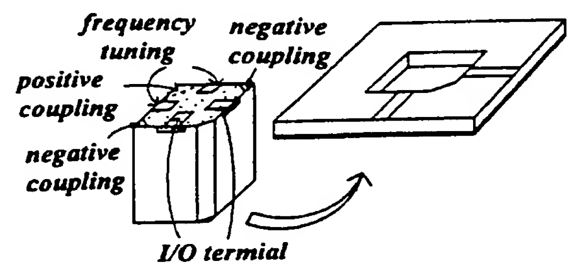

Fig. 14 Configuration of a two-stage bandpass filter.

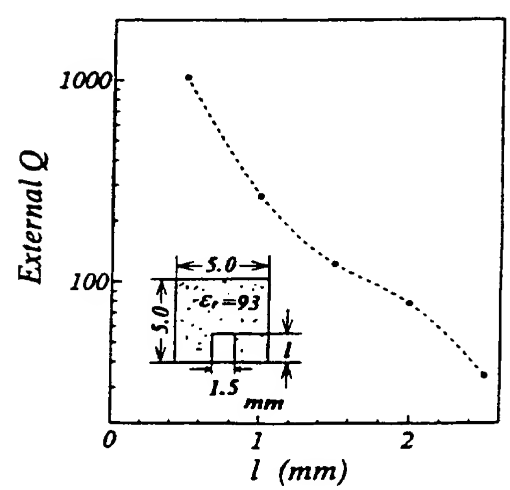

Fig. 15 External Q versus the length of exciting electrode. ($c/2=10$ mm)

shown in Fig. 14. This is no other than the method for a 2-stage bandpass filter. The resonator block is inserted into the square hole of the printed circuit board (PCB), the upper surfaces of both being leveled. The two rectangular electrodes are soldered to the micros-

trip line, and side walls of the resonator are also connected to the ground plane of the PCB.

Figure 15 shows an experimental result for the external Q of each resonant mode according to the dimension of the electrode. Since the coupling is capacitive, the area of the electrode is directly proportionate to the coupling factor $1/Q_e$.

5. Unloaded Q of the Resonator

The unladed Q of a dielectric resonator is determined by the conductor, dielectric and radiation loss. That for a resonator of square cross section is given by [8]

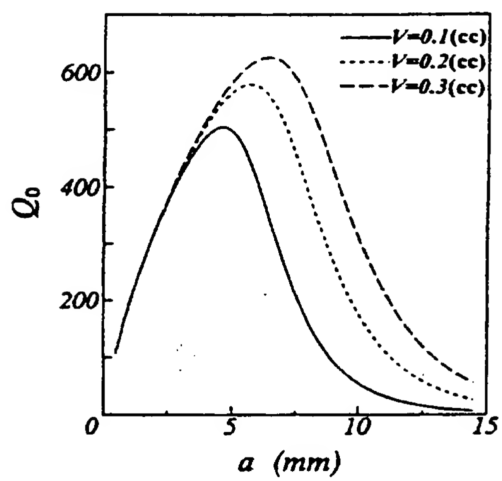

Fig. 16 Theoretical value of unloaded Q versus the cross-sectional dimension. (Volume is kept constant. $A=0.0002$/GHz $\varepsilon_r=93$. Conductivity is taken half of bulk silver)

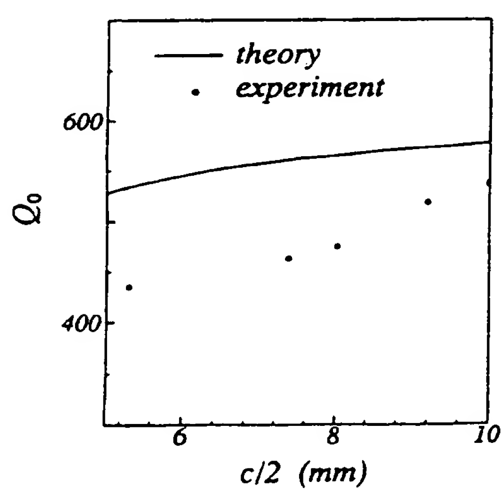

Fig. 17 Unloaded Q versus the length of the resonator. ($a=5$ mm)

$$\frac{1}{Q_0}=\Delta\frac{a^2+c^2+\frac{2}{c}(a^3+c^3)}{a(a^2+c^2)}+Af_0+\frac{32}{3\pi^2}\frac{a^2}{c\lambda_0\varepsilon_r}. \tag{14}$$

where Δ is the skin depth of the silver coating, A is a constant proportionate to the loss of the dielectric, ε_r the relative permittivity, λ_0 the free space wavelength, and f_0 the resonant frequency.

If one keeps the volume of the resonator constant and changes the cross section, Q_0 takes a maximum as shown in Fig. 16. This is because the conductor loss is larger for a smaller cross section, while the radiation loss becomes larger for a greater cross section.

Figure 17 shows the experimental and theoretical values for Q_0 versus the length of the resonator with the dimension of the cross section constant. The difference between them is considered due to the under-estimation of the radiation loss in the theory, since the third term in Eq. (14) which corresponds to the radiation loss takes only the contribution of magnetic current at the end face into account. In fact, we can not neglect the effective electric current due to the magnetic field leaking from the end face.

The conductive loss of the resonator is induced by the surface magnetic field on the surrounding metal coating. Since the field dose not change in first approximation by adding a perturbation like Fig. 3 or Fig. 9 (a), the unloaded Q of a perturbed resonator should remain nearly the same.

6. Two-Stage Bandpass Filter

A two-stage BPF is designed with the help of the insertion loss method. The normalized reactances of the maximally-flat prototype lowpass filter is given as

$$g_0=g_3=1,\ g_1=g_2=1.414 \tag{15}$$

If we choose the center frequency $f_0=3.3$ GHz and bandwidth $B=70$ MHz, the coupling constant is

$$k=\frac{1}{\sqrt{g_1g_2}}\frac{B}{f_0}=0.015 \tag{16}$$

Referring to Fig. 6, one takes the corner cut depth Δd as 0.63 mm. As for the external Q,

$$Q_e=\frac{g_0g_1f_0}{B}=67 \tag{17}$$

should be chosen, resulting in the exciting electrodes of 1.9×1.5 mm by use of Fig. 15. The other metal patterns for tuning described in Sect. 4(a) and (b) are also prepared on the open end face. The filter is attatched to a PCB as shown in Fig. 14

The insertion loss in the pass band can be calculated as

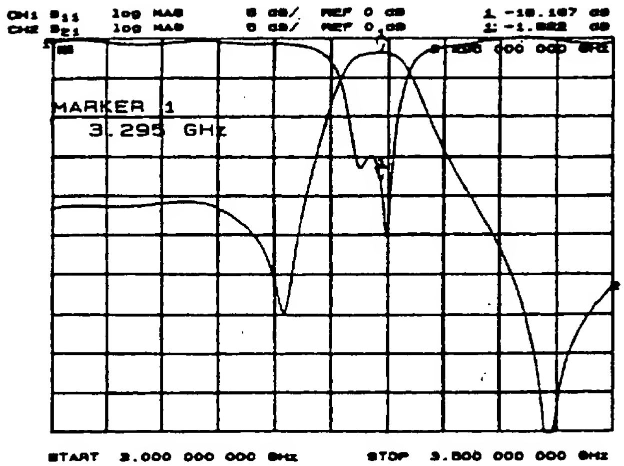

Fig. 18 The transmission and reflection characteristics of a fabricated BPF. (The transmission loss 0.5 dB of the external microstrip line should be excluded)

$$L=4.34\frac{f_0}{B}\sum_{i=1}^{2}\frac{g_i}{Q_{0i}}=1.11\,dB, \qquad (18)$$

where Q_{0i} was read as 520 from Fig. 17

The experimental result is depicted in Fig. 18. Although unexpected notches appear on the both sides of the pass band, the bandwidth and the insertion loss are almost as was designed. The notches are usually desirable because the sharper shoulder response of transmission characteristic improves the out-of-band rejection. They are supposed to be originated from the interference between the double signal paths i.e. the mutual inductive coupling of the two resonant modes and the mutual capacitive coupling of I/O terminals in Fig. 14. We are investigating how to design the notch characteristics including their spacing and depth.

7. Conclusion

A dielectric rectangular waveguide of square cross section has two fundamental modes that are degenerate and spatially orthogonal. We have studied the $\lambda/4$ resonator made of TE_{10} and TE_{01} mode in the waveguide. First, a mutual coupling has been realized via a proper perturbation such as a corner cut or metal patterns on the end face of the resonator. It is controlled by the dimension of those perturbations, which is used for design of the bandwidth of a BPF. Secondly, it was elucidated that the external Q and the resonant frequency can be adjusted by adequate metal electrodes on the end face. The unloaded Q was also measured and shown to have a maximum for the aspect ratio of the cross section and the length of the resonator.

Lastly, we have designed a two-stage bandpass filter on basis of the data above mentioned. The fabricated filter has shown a good coincidence with the designed values, including the center frequency, bandwidth and the attenuation in the passband. The structure of the proposed filter is one of the simplest and also tuning procedure is one of the easiest in terms of internal and external coupling or the center frequency. Extension to the more multiple coupling of resonant modes will be possible.

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4 5. Oktober 2004
~~16. August 2004~~ (Bescheid weg?) OK

Application No.	Ref.	Date
00 953 537.8 - 2220	UH 4-15400.7	05.07.2004
Applicant: Kabushiki Kaisha Tokin		

Communication pursuant to Article 96(2) EPC

The examination of the above-identified application has revealed that it does not meet the requirements of the European Patent Convention for the reasons enclosed herewith. If the deficiencies indicated are not rectified the application may be refused pursuant to Article 97(1) EPC.

You are invited to file your observations and insofar as the deficiencies are such as to be rectifiable, to correct the indicated deficiencies within a period

of 4 months

from the notification of this communication, this period being computed in accordance with Rules 78(2) and 83(2) and (4) EPC.

One set of amendments to the description, claims and drawings is to be filed within the said period on separate sheets (Rule 36(1) EPC).

Failure to comply with this invitation in due time will result in the application being deemed to be withdrawn (Article 96(3) EPC).

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VAN DER PEET H.
Primary Examiner
for the Examining Division

Enclosure(s): 2 page/s reasons (Form 2906)

Registered Letter
EPO Form 2001 07.02CSX

The examination is being carried out on the **following application documents**:
Description, Pages 1-31 as originally filed
Claims, Numbers 1-16 as originally filed
Drawings, Sheets 1/19-19/19 as originally filed

1. In order to facilitate easy reference the documents cited in the ISR are numbered *seriatim* (D1 to D7), whereas the document cited in the Supplementary ESR is referenced D8.

2. The present invention has two objects, to wit: a) reduction of the number dielectric resonators in dielectric filters thus reducing cost and size and b) provision of a very small dielectric resonator with simple composition and triple mode resonance (cf. description bridging pages 4 and 5).
The application contains two independent claims, i.e. claims 1 and 6. Claim 1 has the broadest scope, whereas claim 6 is more restricted by the reference to a specific mode. The number of claims must therefore be reduced to one (Rule 29(2)).

3. The claims are of the apparatus category (dielectric resonator, dielectric filter) couched in method wording. E.g. in claim 1 the modes are coupled b "chamfering" a ridge portion of the dielectric block, whereas this is normally in an apparatus claim expressed as "the modes are coupled by a chamfered ridge portion of the dielectric block". A similar observation holds for claim 2 ("disposing") and *mutatis mutandis* for all other claims (Article 84 EPC, clarity).

4. Claim 4 alludes to "a composition in which resonant frequency of each resonance and an amount of coupling between resonators is adjustable". It is unclear the composition of what renders the frequency adjustable. It is moreover unclear how the composition of any item could render the frequency adjustable. Claim 4 formulates a mere *desideratum* but fails to indicate technical features with which the *desideratum* is achieved (Rule 29(1)).

5. Claim 8 alludes to surfaces "sharing a point" of a dielectric block, whereas according to claim 9 these surfaces are "sharing an apex". This renders claims 8 and 9 unclear. It is moreover a mystery how a surface ratio of 200% can be realised.
Claim 11 refers to an apex which has no antecedent in claim 8. It is also unclear how a single feeding probe can be provided on surfaces B and C at the same time.

The directions p and p' in claim 12 are wholly indeterminate. Claim 13 is incomprehensible because the attenuation pole referred to has no antecedent.

6. Having regard to the dielectric resonator known from figure 4 of document D1 the present application does not seem to contain patentable subject matter. In the event the applicant wishes to pursue the present application he is invited to file a single new independent claim overcoming the above formulated clarity objections. The claim must be correctly delimited in respect of document D1 (Rule 29(1)) and preferably supported by a convincing argument concerning inventive step. The claims must contain reference numerals (Rule 29(7)). Document D1 must be acknowledged in the description (Rule 27(1)b).
Attention is drawn to the proscription of Article 123(2) EPC.